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Technical Note 1975-65

S. Seneff

A Real-Time
Digital Telephone Simulation
on the Lincoln
Digital Voice Terminal

30 December 1975

Prepared for the Defense Communications Agency
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MASSACHUSETTS INSTITUTE OF TECHNOLOGY

LEXINGTON, MASSACHUSETTS

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This technical report has been reviewed and is approved for publication.

FOR THE COMMANDER

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LINCOLN LABORATORY

A REAL-TIME DIGITAL TELEPHONE SIMULATION ON THE LINCOLN DIGITAL VOICE TERMINAL

S. SENEFF
Group 24

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ABSTRACT

A telephone channel simulator has been implemented on the Lincoln Digital Voice Terminal (LDVT). This technical note first reviews the various distortions that occur in telephone channels and then describes the implementation. This work was undertaken at the suggestion of Mr. Ronald Sonderegger of the Defense Communications Engineering Center who observed that a real-time simulation on the LDVT should be far less costly than previous implementations. The results obtained indicate that the potential of the LDVT goes beyond the original intent of its use as a speech digitizer.

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INTRODUCTION

There has been growing interest recently in the feasibility of interconnecting the telephone system with the various digital vocoders that are currently being developed. The obvious possible uses of the telephone system are: 1) to transmit speech from a local source to the input of a vocoder, and 2) to transmit the bits from a vocoder analyzer to a vocoder synthesizer. Each of these possible functions has a set of problems associated with it, due to the various distortions and interferences in a telephone line. For example, it is generally known that pitch detection algorithms, upon which many vocoders depend, break down when the input is telephone speech. When bits are transmitted over a telephone line, they must be converted to a type of signal which can be sent with a minimal number of errors. This task is assumed by a modem, of which there are many varieties on the market. What makes this task difficult is the fact that the telephone channel makes no attempt to correct for phase distortion, since the human ear is presumed to be insensitive to phase delay.

By simulating the telephone channel's characteristic distortion and interference in real time in a digital computer, it should be possible to investigate the reasons for the problems encountered when actual telephone lines are used, both as inputs to a vocoder and as transmission media for bits. Using the simulation, one can hopefully develop more robust modems and vocoders, and one can determine how much of a given type of interference is required before a given system breaks down.

This report begins with a discussion of how the telephone system is constructed. Each of the various types of interference, Gaussian noise, frequency attenuation and phase delay, quadrature distortion, nonlinear distortion, echo, and cross talk, is then discussed in some detail. Following that, a description is given of how the measurements were obtained for the simulation. Finally, the details of the actual simulation are described fully, including an indication of those areas in which the simulation is a simplification of the actual situation.

I. Overview-Description of a Telephone Channel

In a typical telephone channel[1], the signal is first filtered, then modulated up to some carrier frequency, transmitted with multiplexing through a series of cables and repeaters and finally filtered and demodulated at the receiver. Various distortions are introduced into the signal at all parts of the system. The filters pass only energy between 300 and 3000 Hz, and introduce both amplitude and phase distortion in the pass band. The modulation demodulation process introduces quadrature distortion whenever there is an offset between the modulating frequency and the demodulating frequency. Cross talk is a result of both frequency and space division multiplexing, whenever the filtering and/or shielding is inadequate. In travelling down the cable, the signal becomes attenuated and therefore the repeaters are necessary to amplify it at various points in the transmission lines. The repeaters are a source of both Gaussian noise and nonlinear distortion. When a signal is switched from one carrier to another there is introduced a sudden change in the phase relationship, as well as transient (impulse) noise. Another source of impulse noise is lightning and corona type discharges.

II. SSB Modulation

Typically, voice signals are frequency multiplexed over a telephone line using single sideband suppressed carrier (SSBSC) amplitude modulation,[2,3] as shown in Figure 1. In order to set the stage for our subsequent discussion of phase distortion, SSBSC is reviewed in this section.

The signal is first bandpass filtered from 300 to 3000 Hz and then modulated up by a carrier cosine whose frequency ω_c is a multiple of 4000 Hz. The resulting signal is then bandpass filtered to eliminate any energy below ω_c. Since the original signal was real, this entails no loss of information. The signal is then frequency multiplexed with several other signals at different carrier frequencies (also multiples of 4 kHz) and is detected at the receiver by means of a bandpass filter from $\pm\,\omega_c$ to $\pm(\omega_c$+4000 Hz), followed by demodulation using a cosine also at frequency ω_c. However, no attempt is made to assure that the modulating and demodulating cosine waves are in phase with each other. In fact the modulating and demodulating frequencies are rarely exactly the same, and any slight difference in frequency will cause a gradual drift into and out of phase.

When the modulating and demodulating cosines are exactly 90° out of phase the resulting signal which is received is the quadrature component, or Hilbert Transform,[4] of the original signal. The quadrature component can be understood from several different viewpoints. If one represents the original signal f(t) in terms of its Fourier expansion, i.e., as a sum of cosines at different frequencies with differing amplitudes and phases, then the Hilbert Transform $\hat{f}(t)$ could be constructed by converting all of the cosines to sines and adding them up. An impulse, for example, is a sum of cosines all in phase at $\theta = 0$,

at that point in time when the impulse occurred. If all of these cosines were converted to sines, every function would be zero at the time of the impulse, all would be negative immediately before it, and all would be positive immediately after. Some distance away on either side, the sines would be at random phase with respect to one another, and hence the net sum at any point would be Ø. The result is the function shown in Figure 2.

One way to acquire the Hilbert Transform is to convolve the signal with the function of Figure 2. Another way is to take the Fourier transform, multiply positive frequency by -j and negative frequency by +j and take the inverse Fourier transform. It is clear that such a filter would convert a cosine at any frequency, ω, into a sine, since the transform of a cosine is two positive real impulses at $\pm\omega$ and the transform of a sine is a negative imaginary impulse at $+\omega$, and a positive imaginary impulse at $-\omega \cdot ((e^{j\omega} - e^{-j\omega})/2j = \sin\omega)$.

Supposing one had available a complex signal z(t) whose spectrum was zero for negative frequency and identical to the spectrum of a given real signal f(t) for positive frequency. According to the previous discussion, such a function could be generated as follows:

$$z(t) = 1/2\ (f(t) + j\ \hat{f}(t)) \quad \text{(Figure 3)}$$

If one were to multiply z(t) by $e^{j\omega_c t}$, the transform of the resulting function would be simply $Z(\omega)$ translated up by a frequency ω_c, i.e., $Z(\omega - \omega_c)$. Now, were one to take the real part of this signal and determine its transform the result would be the even part of $Z(\omega - \omega_c)$ which (Figure 4) is clearly the same as what would be obtained by multiplying f(t) by $\cos\omega_c t$ and filtering out energy below ω_c. Remembering that the even part of a transform is the transform

of the real part of the function, we have:

$$f_s(t) = 2\,\mathrm{Re}\,[z(t)\, e^{j\omega_c t}]$$

$$= \mathrm{Re}\,[(f(t) + j\,\hat{f}(t))\, e^{j\omega_c t}]$$

$$= [f(t) \cos \omega_c t - \hat{f}(t) \sin \omega_c t].$$

Now it should be straightforward to explain the effect of a phase difference between the modulating and demodulating carriers.

$$f_o(t) = [f(t) \cos \omega_c t - \hat{f}(t) \sin \omega_c t)] \cos (\omega_c t + \phi) \quad \text{(Low Pass)}$$

$$= f(t) \cos \omega_c t \cos (\omega_c t + \phi) - \hat{f}(t) \sin \omega_c t \cos (\omega_c t + \phi) \quad \text{(Low Pass)}$$

$$= f(t) \cos \omega_c t \cos \omega_c t \cos \phi - f(t) \cos \omega_c t \sin \omega_c t \sin \phi$$

$$- \hat{f}(t) \sin \omega_c t \cos \omega_c t \cos \phi + \hat{f}(t) \sin \omega_c t \sin \omega_c t \sin \phi \quad \text{(Low Pass)}$$

Recalling that

$$\cos^2 \theta = 1/2\ (1 + \cos 2\theta)$$

$$\sin^2 \theta = 1/2\ (1 - \cos 2\theta)$$

$$\sin \theta \cos \theta = 1/2\ (\sin 2\theta)$$

and that the $\sin 2\theta$, $\cos 2\theta$ terms will all be clobbered by low pass filtering at the output, we have finally

$$f_o(t) = 1/2\ (f(t) \cos \phi + \hat{f}(t) \sin \phi)$$

When $\phi = 90°$, $f_o(t) = 1/2\ \hat{f}(t)$, which is to say that only the quadrature term is received.

III. Disturbances in the Transmission System

A. Quadrature Distortion

Armed with a knowledge of the mathematics of SSB carrier systems, it is now simple to understand the various phase distortions that occur in the telephone systems. If there is a frequency difference $\Delta\omega$, between the modulating and demodulating waves, we would have:

$$f_o(t) = (f(t) \cos \omega_c t - \hat{f}(t) \sin \omega_c t) \cos ((\omega_c + \Delta\omega)t) \text{ (Low Pass)}$$

The contribution to the phase difference, ϕ, by the frequency offset, ω, is a function of time:

$$\phi_o = (\Delta\omega)\, t$$

which accounts for a slow drift into and out of phase.

Phase jitter is the term used to describe the interference of a sinusoidal noise (frequently 60 cycles) in the phase of a carrier. Sixty cycle interference shows up everywhere in the system but in general is harmless because of the 300 Hz lower limit of the filters. However, 60 cycle interference in the generation of a cosine causes a phase distortion of the cosine waveform which appears as a very low frequency modulation (Figure 5). The phase jitter introduces an additive component to the phase which can be expressed as

$$\phi_j = A_j \cos \omega_j t$$

where A_j is the amplitude of the jitter in degrees and ω_j is usually 60 cycles (50 cycles in Europe).

The final phase effect is phase hits, or sudden large shifts in phase, which occur as a result of switching two carrier supplies not in phase. Sometimes a phase coherent detector corrects the situation some time later. On other occasions the original phase relationship is never restored.

The net phase relationship is a combination of the three effects:

$$\phi = \phi_o + \phi_j + \phi_h$$

and the final output signal is

$$f_o(t) = f(t) \cos \phi + \hat{f}(t) \sin \phi .$$

B. Filtering

A major part of the distortion in a telephone signal has to do with the frequency response and phase delay characteristics of the filter used to assure that the signal remains in band. Theoretically, in multiplexing at 4000 Hz intervals, one could pass all frequencies from 0 to 4000 Hz and avoid cross talk. In practice, filters never cut off sharply, and so a healthy compromise is to pass the signal from about 300 to 3000 Hz. In addition, there tends to be considerable noise interference at low frequencies (for example, 60 cycle hum) so that it is advantageous from a signal to noise ratio argument to remove low frequencies. The characteristics of the filter in the pass band represent a tradeoff between cost and response. In particular not much attention was paid to phase delay distortion, as the ear is insensitive to phase.

C. Nonlinear Distortion

In the process of travelling down a coaxial cable, a signal's energy is gradually attenuated due to losses in the transmission line and therefore it is necessary to amplify the signal at certain points in the transmission path to assure an adequate level at the receiver. The devices which achieve the amplification are called repeaters, and are generally made up of two-port amplifiers. The voltage transfer characteristics of a generalized two-port are shown in Figure 6. According to the Fourier series expansion, the transfer characteristics can be expressed as:

$$e_o = a_1 e_i + a_2 e_i^2 + a_3 e_i^3 + \ldots$$

where a_1 is the gain and the other a's are the nonlinear distortion coefficients.

The sources of the nonlinear distortions are several, and much has been written on the subject of repeater design (see, for example, Bell, pp. 396-421, Ref. 1). Transistors have an inherent nonlinearity dependent on the source and load impedances and the selected d-c bias conditions. Negative feedback in the amplifier complicates the nonlinear relationship. The nonlinearities in a repeater are, in general, frequency dependent and amplitude dependent, and the calculation of the precise response for a given system design is quite complicated.

D. Gaussian and Impulse Noise

Gaussian noise is an unavoidable side product of networks and devices. Two common types of noise in a circuit are thermal noise and shot noise.[5] Arguing from the central limit theorem, both can be assumed to be Gaussian and white at the source. Thermal noise is present in any conductor, for example, a resistor, and is due to the thermal interaction between free electrons and vibrating ions. Its available power is directly proportional to the product of the bandwidth of the system and the temperature of the source. Shot noise is due to the discrete nature of electron flow and is present in most active devices (transistors and diodes). Its amplitude is proportional to the square root of the current and therefore is dependent on signal level.

The components of the repeaters are a major source of the Gaussian noise. By the time the noise reaches the receiver, it is no longer white, because it has been shaped by the filters at demodulation.

In addition to Gaussian noise there is occasionally a burst of noise on a line whose amplitude far exceeds the average noise level of the system.

Such noise is referred to as "impulse hits" and is generally caused by switching transients in central offices or from corona type discharges (electrical discharges in the air surrounding a high potential line) that occur along a repeated line.

E. Echo and Cross Talk

To be complete in a report on telephone channel characteristics one would have to include at least a brief discussion of echo and cross talk, even though these two effects were not included in the digital simulation. An echo may be produced whenever there is an impedance discontinuity. It is most commonly caused by a return of a talker's signal through the channel to which he is listening. For short distances the effect is indistinguishable from side tone and is therefore not annoying. Echo suppressor circuitry has been added to long distance lines to attenuate the signal on the return path when the received signal amplitude is high. The result is that when both people talk only one is heard, and sometimes the beginnings of words are clipped. There is, however, a "tone disabler" section of the echo suppressor circuit which permits the mechanism to be shut off when data is being transmitted.

Cross talk is caused by coupling losses between two active circuits. Transmittance cross talk occurs because of inadequate design of modulators and filters in frequency multiplexing. Coupling cross talk is caused by electromagnetic coupling between two physically isolated circuits. The result, of course. is that the listener hears another conversation in the background, which may or may not be intelligible.

IV. Measuring the Telephone Channel

Before the telephone channel could be simulated, it was necessary to make extensive measurements on thousands of lines in order to determine concrete values to be used in the various components of the simulation. From the study data set, histograms were compiled for all of the various parameters, such as frequency offset and level of Gaussian noise, for Continental U.S. (Conus) and European voice and data grade lines. Table 1 is a chart of the resulting numbers for the various degradations (excepting filter frequency response).[6] The term "mid" is used to refer to the 50th percentile on the histogram and "poor" to refer to the 90th percentile; i.e., 90% of the Conus voice grade channels measured were better than "Conus poor voice".

Gaussian noise and nonlinear distortion were measured in units of dBmc, meaning dB's relative to 1 milliwatt, using a special C-message weighting curve[7](Figure 7). This frequency weighting curve was determined empirically by means of several subjective listener tests and is intended to reflect the amount of listener annoyance for noises at different frequencies. The unit of measurement used for impulse noise is dBrnc, where rn stands for reference noise, which is -90 dBm, or 10^{-12} watts at 1000 Hz.

The device used by the telephone company to measure Gaussian noise has built in, in addition to the C-message weighting curve, a mechanism for attenuating impulse noise of a sufficiently short duration. The argument again is one of ear sensitivity, as the human ear does not fully perceive the power in noises of duration less than 200 msec.

The telephone company has available another device, called an impulse

counter,[8] for the purposes of measuring the amount of impulse noise present, as such noise is far more destructive than Gaussian noise in introducing errors in the transmission of bits through a modem and a channel. The counter consists of a weighting network, a rectifier, a threshold detector, and a counter of events above threshold. Several impulse counters can be used simultaneously at different thresholds to obtain information about the distribution of the magnitudes of the impulses.

To measure the nonlinear distortion,[9] a pure tone at a fixed level is transmitted, and the amount of energy received at the 2nd and 3rd harmonics is measured. From these values, one can deduce the coefficients a_2 and a_3 of the squared and cubed terms:

$$\cos^2\theta = 1/2\ (1 + \underline{\cos 2\theta})$$

$$\cos^3\theta = \cos\theta \cos 2\theta = 1/2\ (\cos\theta + \cos\theta \cos 2\theta)$$

$$= 1/2 \cos\theta + 1/4\ (\underline{\cos 3\theta} + \cos\theta)$$

Because of this relationship between nonlinearities and harmonics, nonlinear distortion is often referred to as harmonic distortion.

V. Simplifications Used in the Simulation

The parameters of the simulation were set so as to emulate, as closely as possible, the results found from the statistical study. In the case of certain measures, such as frequency offset and phase jitter, this was a straightforward process. However, in the simulation of Gaussian noise, impulse hits, phase hits, and nonlinear distortion, certain (sometimes gross) simplifications were used.

The numbers used for Gaussian noise, impulse noise, and nonlinear distortion in the simulation were determined empirically by adjusting constants until the desired value was obtained at the output by the appropriate measuring device (described in section IV).

Gaussian white noise generated at various devices in the transmission line is no longer white by the time it reaches the receiver because of the filter transfer characteristics. In the simulation, Gaussian noise is added at the input, so that it will be shaped by the filter of the system. The level of the Gaussian noise in the telephone system is dependent upon the amplitude of the input signal, whereas in the simulation noise is kept at a fixed level.

Impulse hits and phase hits are simulated to occur at fixed intervals and at fixed duration and level, so as to correspond in frequency of occurrence and average amplitude to the given telephone channel characteristics.

Nonlinear distortion is simulated by squaring and cubing the final output of the system. The coefficients a_2 and a_3 remain fixed and are equal, even though in the actual telephone system these two coefficients would in general be dependent on both amplitude and frequency. The appropriate gain for the squared and cubed terms was of course determined so as to match the desired meter measurement. This was done by sending a pure tone at 700 Hz through the digital distorter and measuring the output of a notched filter (to remove 700 Hz) in dBrnc.

VI. Implementation on the LDVT

The Lincoln Digital Voice Terminal (LDVT) is a small high speed computer which has proven capable of realizing in real-time a variety of algorithms

for low bit rate digital speech transmission[10] (from 2400 bits/sec to 16000 bits/sec). The computer consists of a 1024 word program memory, Mp, with a cycle time of 55 nsec, a 512 word data memory, Md, and a 2048 word outboard memory, Mx, an input/output device from which data is accessible in a few instruction cycles. The LDVT has a very simple instruction set and is quite convenient to program.

The telephone channel simulator represents the first attempt to use the LDVT for something other than a vocoder. A block diagram of the complete simualtion is given in Figure 8. The input speech is low pass filtered to remove energy above 5 kHz and sampled at 10 kHz using a 12 bit A/D converter. Impulse and Gaussian noise are generated digitally and added to the input samples.

The frequency response and delay characteristics of the telephone filter and the Hilbert transform to obtain the quadrature component are both realized by means of high speed convolution techniques.[11] After a 256 point FFT of the input samples (256 points is adequate as long as the combined impulse response of the filter and Hilbert transform is less than 12.8 msec), negative frequency is zeroed out and positive frequency is complex multiplied by the FFT of the desired impulse response. A 256 point IFFT now yields the filtered speech $f(t)$ in the real data buffer and the filtered quadrature component, $\hat{f}(t)$, in the imaginary data buffer (refer to the previous discussion on SSB modulation). Only the second half of the data is good, because of circular convolution, and therefore an overlap of 128 samples is necessary between frames.

The phase, ϕ , is computed as the sum of the contributions of phase jitter, frequency offset, and phase hits. The output of the "modulator demodulator"

complex is then computed as:

$$g(n) = f(n)\cos\phi + \hat{f}(n)\sin\phi .$$

The final step in the simulation is to subtract from g(n) the nonlinear distortion term:

$$\hat{s}(n) = g(n) - D(g^2(n) + g^3(n)).$$

There is a special buffer in Md (data memory) which is set aside as the parameters of the system (excepting filter frequency response) and which can be modified under user control. Figure 9 is a copy of that section of the program, including the appropriate settings for Conus poor voice. The same mnemonics are used in Figure 8, where it should be clear how the various parameters are being used.

Table 2 gives the values used for the various parameters in the simulation of the eight channels. The values for the parameters FRQOFF, PKTOPK, JITTER, PHSHIT, PKTFRC and OFFFRC are simply copied over from Table 1. The values for PHSRTE and IMPRTE are determined as the reciprocal of #/15 min, and converted from units of minutes to units of .01 sec. The parameters PHSDUR and IMPDUR are left unspecified in Table 1, and are set arbitrarily to .01 sec in Table 2 for all channels. The parameters GAUSML, HARM, and IMPAMP are dimensionless fractions determined empirically so as to read the correct meter reading at the output in units of dBmc or dBrnc.

The only other information that is needed to simulate each of the channels is the frequency domain characteristics of the telephone filter. Tables were given for each channel of 256 real and 256 imaginary frequency components of the desired filter, spaced by 20 Hz, spanning 0 to 5 kHz. Since the DVT implementation used a 256 rather than 512 point FFT, alternate samples

were ignored in entering the tables into the computer. It should be noted here that it would have been essentially impossible to implement a 512 point FFT in the DVT due to its current memory limitations, but that the 12.8 msec window appears to be adequate for the filters simulated.

The main body of the program is the FFT-IFFT computation which requires several buffers in Mx (outboard memory), two 128 point buffers in Md and a complex interchange of data between Mx and Md. The program takes advantage of certain characteristics of the data to reduce both time and memory requirements. By keeping the FFT size down to 128, one avoids the necessity of referencing Mx in the inner loop which would greatly increase the time required. Therefore the forward FFT is realized (since the input data is real) by packing even numbered samples in the real Md buffer, MDREAL, and odd numbered samples in the imaginary Md buffer, MDIMAG, and doing a 128 point FFT followed by even odd separation and a final stage of the 256 point FFT. The first stage of the inverse FFT is nonexistent, since the second half of the data is zero, and therefore it can be conveniently split into two 128 point FFT's, with the first one yielding the even numbered output samples and the second one (by beginning each stage with the coefficient index set at half its increment instead of at zero) yielding the odd numbered output samples.

Because of this even numbered odd numbered sorting at both the output and the input, it is convenient to store all buffers of speech in Mx in this peculiar fashion of even numbered samples in one buffer and odd numbered samples in another. The buffers required in Mx are listed in Table 2. The even num-

bered input samples are stored in EVIO and the odd numbered in ODIO. At each A/D interrupt the next sample is fetched out of either EVIO or ODIO alternately and sent to the D/A converters and the new sample from the A/D converter is written over the same place in Mx. At the beginning of a new frame, the most recent 128 samples from the A/D converter have filled up EVIO and ODIO, and so these buffers now become EVNEW and ODNEW, respectively. At the same time, the former EVNEW, ODNEW become EVOLD, ODOLD, and the former EVOLD, ODOLD whose contents had been written over by $\hat{s}(n)$ in the course of the previous frame, become EVIO and ODIO, ready to be sent to the D/A converter.

The first step in a new frame is to fill MDREAL with 128 even numbered new samples from EVOLD/EVNEW and to fill MDIMAG with the odd numbered samples from ODOLD/ODNEW. Now a 128 point normal to bit reversed decimation in frequency FFT followed by bit reversed even odd separation and a final stage of the 256 point FFT can all be realized in place in Md. The resulting data (the first 128 samples of the 256 point FFT) is then complex multiplied by the bit reversed filter coefficients (which have been stored in Mx). The resulting filtered spectrum is then saved in Mx: the real data is stored over ODOLD/EVOLD and the imaginary data is saved in a 128 point buffer, MXIMAG. A 128 point IFFT at this point, of the data in Md, yields the even numbered f(n) samples in MDREAL and the even numbered $\hat{f}(n)$ (quadrature component) samples in MDIMAG.

Only the 2nd half of the data is good, because of circular convolution, and these 64 even numbered f(n) and even numbered $\hat{f}(n)$ are saved in Mx. Meanwhile, the original real and imaginary outputs of the forward FFT filtered

spectrum are fetched back from Mx and another 128 point IFFT, beginning each stage with the coefficient index set at half its increment instead of at Ø, yields the odd numbered samples of $f(n)$ and of $\hat{f}(n)$ in Md. Again, the first 64 of each are garbage. The 64 even numbered samples of $f(n)$ and of $\hat{f}(n)$ can now be fetched back from Mx and stored over the 64 garbage points in each of MDREAL and MDIMAG. Now all of the pertinent data resides in Md and all that is left is the introduction of quadrature distortion and nonlinear distortion:

$$g(n) = f(n) \cos \phi + \hat{f}(n) \sin \phi$$

$$\hat{s}(n) = g(n) - D\,(g^2(n) + g^3(n))$$

This can conveniently be done at this time and the resulting $\hat{s}(n)$ evens can be stored over EVOLD and $\hat{s}(n)$, odds, over ODOLD.

When the interrupt routine has been serviced 128 times, resulting in 64 new samples in each of EVIO, and ODIO, it is time to swap buffer pointers again, leaving $\hat{s}(n)$ evens in EVIO and $\hat{s}(n)$ odds in ODIO, as desired, and we have come full circle.

The assurance of adequate accuracy without overflow in the FFT-IFFT complex at first posed some difficulty, as a check for overflow and correction in the inner loop proves very costly in terms of time. The scheme that was finally decided upon was the following:

1. Shift the input buffer of the forward FFT in Md as far left as possible as a block so that the largest sample does not overflow.

2. Scale down the data as a block by 1/2 at each stage of the inner loop of the forward FFT.

3. Scale the output data in Md from the forward FFT as far left as possible.

4a. If the data has been scaled up by more than 2^N, scale it down by 1/2 at each stage of the IFFT and scale it down the remaining amount at the end.

4b. If the data has been scaled up by less than 2^N, scale it down by 1/2 at each stage of the IFFT until such time as no more scaling is needed.

The remainder of the program is reasonably straightforward. Gaussian and impulse noise are added to a new s(n) in the interrupt routine, as soon as it comes in from the A/D converter and before it is sent to Mx. Gaussian noise is simulated by first generating a 9-bit pseudorandom number with a flat distribution. Eight bits are used as a pointer into a table of the midpoints of 256 equal areas over the positive half of a Gaussian distribution, and the ninth bit determines the sign. The numbers in the table were chosen such that $\sigma = .25$, and values, when fetched from the table, are multiplied by the appropriate constant, GAUSML, to obtain the desired σ. Impulse noise is created by adding to each s(n) a signal of height IMPAMP for the duration of time IMPDUR, and at a periodic rate IMPRTE.

The phase, ϕ, and the components ϕ_o, ϕ_h, and ϕ_j, are all stored in the computer as fractions, with 1.0 corresponding to 360°. ϕ_o is updated at each s(n), by adding FRQOFF/10000, since an offset of 1 Hz would be 360° in 10000 samples. ω_j, likewise, is updated by adding JITTER/10000 at each s(n). The cosine of ω_j is multiplied by PKTOPK/360 (expressing degrees peak to peak as a fraction) to obtain ϕ_j. Phase hits are simulated by adding PHSHIT/360 to ϕ

for those samples s(n) during the interval PHSDUR and spaced by the time PHSRTE. The three values cos ω_j, cosϕ and sinϕ are determined by table lookup from Mx using the same first quadrant 64 point cosine table as is used for the FFT's plus an additional interpolation angle of 90°/128 whose cosine and sine are stored in Md. From these it is possible to determine the cosine and sine of any angle from 0-360° to the accuracy of 90/128° using the formulas:

$$\cos(a+b) = \cos a \cos b - \sin a \sin b$$

and

$$\sin(a+b) = \cos a \sin b + \cos b \sin a$$

Harmonic distortion is added just before the final adjustment of the data from block floating to fixed point so as to gain bits in the cubing and squaring of the data. The terms $g^2(n)$ and $g^3(n)$ are added together, multiplied by the parameter HARM, and subtracted from g(n) to obtain $\hat{s}(n)$, which when converted back to fixed point, is the final output of the system.

A somewhat intuitive feel for the effects of phase jitter and frequency offset can be gained by referring to Figure 10. This figure shows a time exposure of the impulse response of the telephone filter for Conus poor voice with a) all of the other distortions removed from the system, b) only phase jitter added and c) only frequency offset added. It can be noted that with a phase jitter interference the samples "jitter" about a focal point, whereas with a frequency offset there is a smooth motion of the wave as it appears to continually pass through and disappear.

VII. DVT Times and Space

Table 3 is a list of the various subroutines in the LDVT in their approximate order of occurrence. The table includes for each subroutine the amount of

program memory required and the amount of time required for its execution. The total time needed per frame is slightly more than half of the time available and the program uses up 77% of Mp.

There are only 512 locations in data memory, Md, of which half is needed for the 128 point FFT real and imaginary buffers. Fortunately, no other large buffers are needed by the program, so that an additional 125 locations were needed for the various parameters, temporaries and variables, leaving one fourth of Md unused.

As for outboard memory, Mx, 256 locations are needed for the filter coefficients, 256 for the Gaussian table, 128 for the cosine tables, and 4 x 128 for the various speech buffers, for a total of 1152 or 56% of Mx (Table 4).

VIII. Summary

A digital telephone channel simulator has been implemented on the LDVT which operates well within the real-time requirements and allows for comfortable margins in each of the various LDVT memories (Mp, Md, Mx). Included in the simulations are Gaussian noise, impulse hits, filter frequency response and delay characteristics, phase jitter, frequency offset, phase hits, and nonlinear distortion. The sources of each of these impairments in the telephone channel and the techniques for measuring them have been discussed, as well as the limitations of the simulation compared to the real situation.

One could easily set up a system to send input speech through a telephone channel simulator in one LDVT into a vocoder (LPC, APC, etc) in another LDVT to investigate the amount and type of degradations caused by tandeming a

telephone line with a vocoder. The telephone simulation has been implemented so as to make it convenient for a user to vary the parameters of the system online, making it possible to determine which aspects of the telephone system are damaging to the vocoder in what way, and then to develop the vocoder to be more robust against these impairments.

Discussions have been evolving recently concerning the possibility of implementing a digital modem on the LDVT. One can imagine at some future time a series of LDVT's representing telephone channels, modems, and vocoders, which can open the door to a wide variety of possible experiments.

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The author would also like to thank Captain Russell Lemon of the National Security Agency for providing the necessary data for the various channels simulated and for many fruitful phone conversations elucidating the information.

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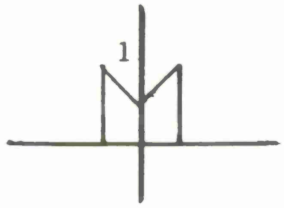
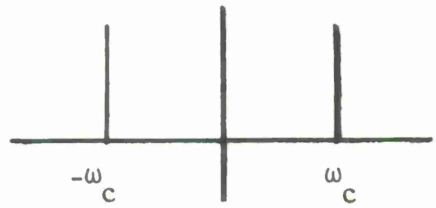
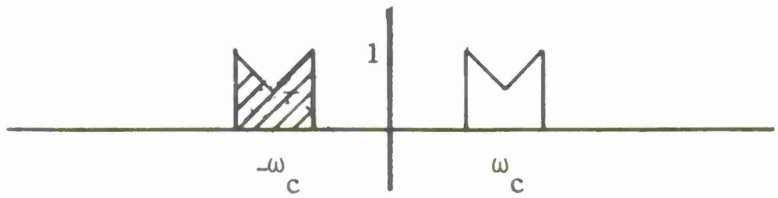


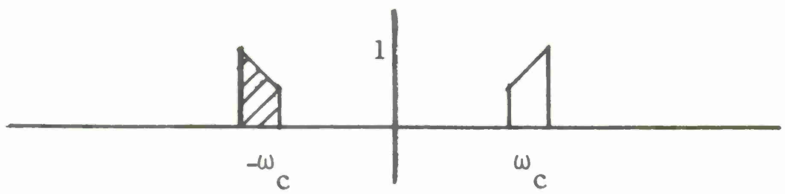


$F(\omega)$ (transform of $f(t)$)

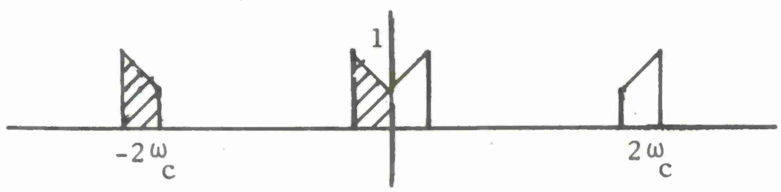
$F_{cos}(\omega)$ (transform of $\cos \omega_c$)



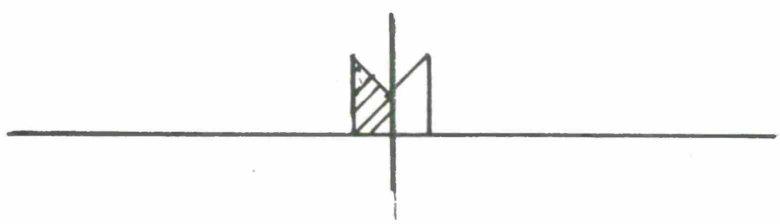
a) Modulated up



b) Band pass filtered to exclude frequencies below ω_c.



c) Modulated down at receiver



d) Low pass filtered to restore $F(\omega)$

Fig. 1. Single side band modulation demodulation technique.

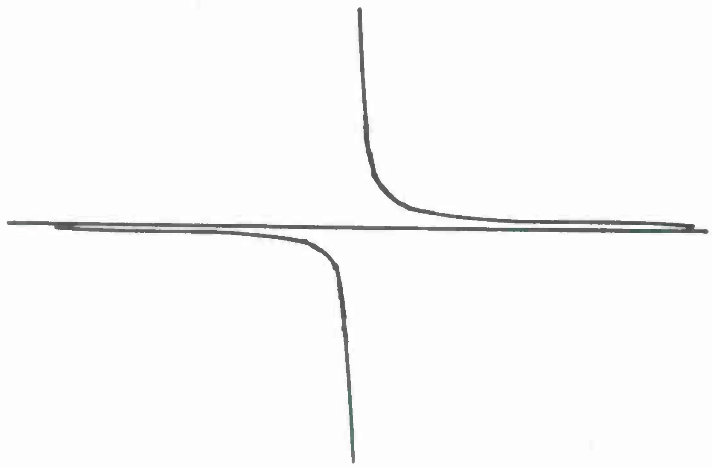
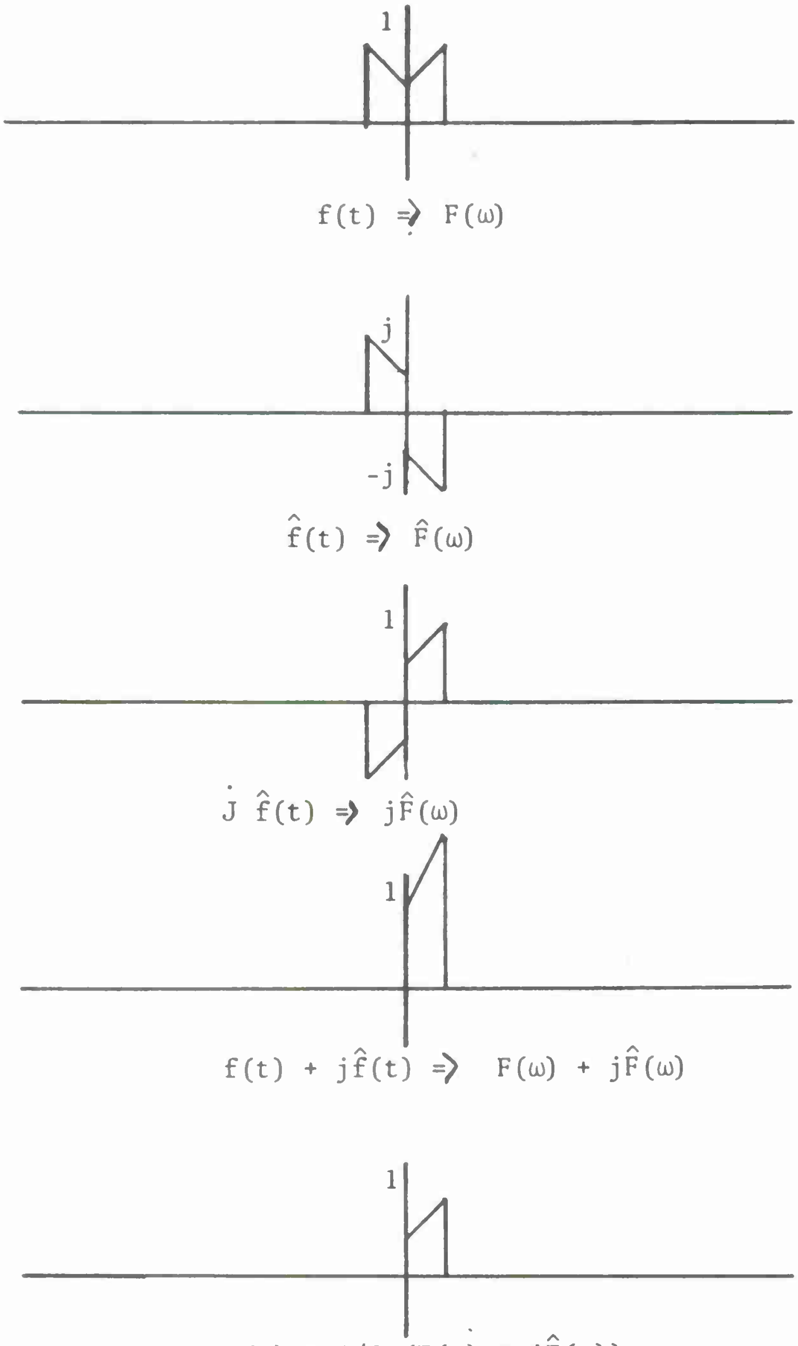


Fig. 2. Hilbert transform of an impulse.

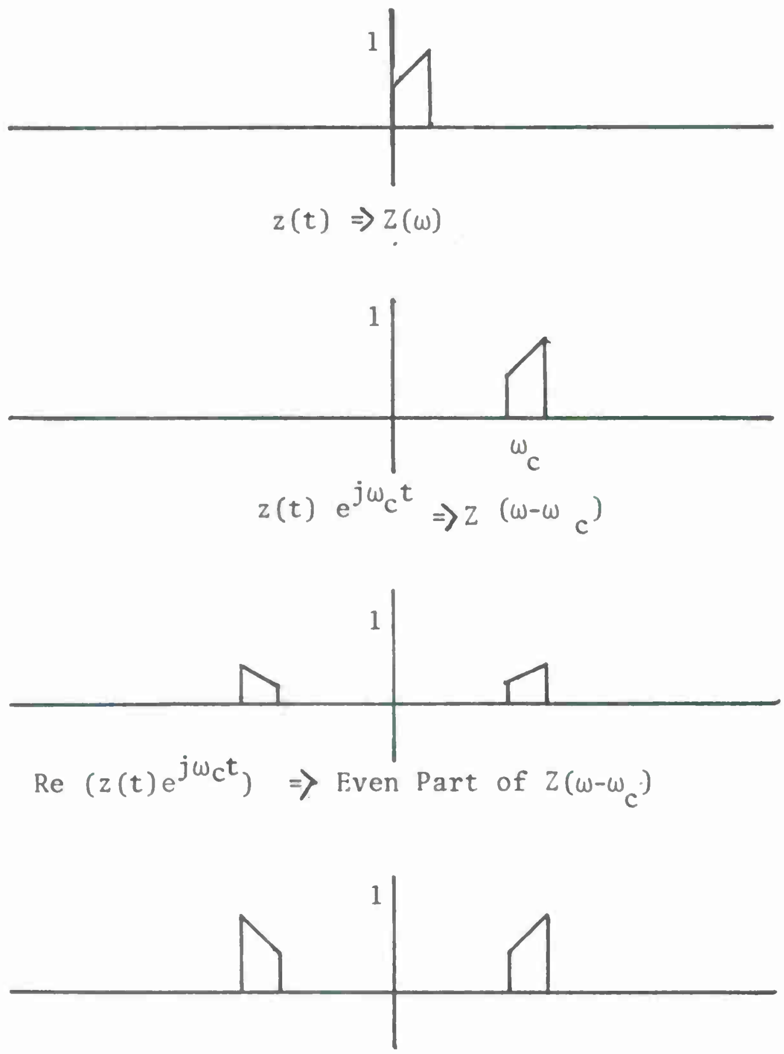
TN-1975-65(3)



$$Z(\omega) = 1/2\ (F(\omega) + j\hat{F}(\omega))$$

$$z(t) = 1/2\ (f(t) + j\hat{f}(t))$$

Fig. 3. Analytic signal z(t) expressed as a function of real signal f(t) and its Hilbert transform f(t). (See Text for Discussion).



$z(t) \Rightarrow Z(\omega)$

$z(t)\ e^{j\omega_c t} \Rightarrow Z\ (\omega-\omega_c)$

$\mathrm{Re}\ (z(t)e^{j\omega_c t}) \Rightarrow$ Even Part of $Z(\omega-\omega_c)$

$f_s(t) \Rightarrow F_s(\omega)$

$f_s(t) = 2\ (\mathrm{Re}\ (z(t)e^{j\omega_c t}))$

Fig. 4. Single side band signal, $f_s(t)$, expressed as function of analytic signal, $z(t)$.

TN-1975-65(5)

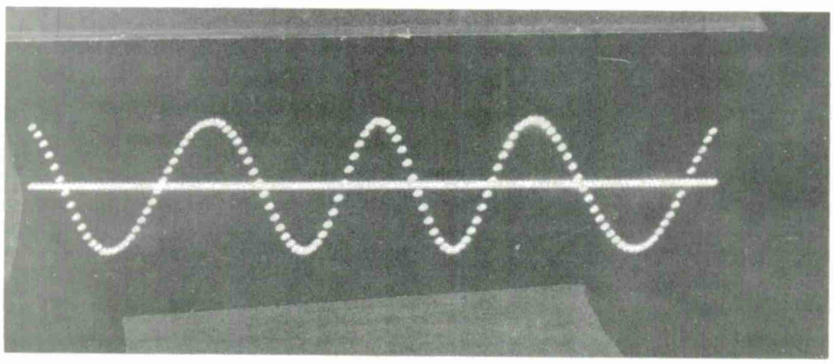


Fig. 5. Cosine wave at 312 cycles/sec distorted by phase jitter at 60 cycles/sec, 30° peak-to-peak.

TN-1975-65(6)

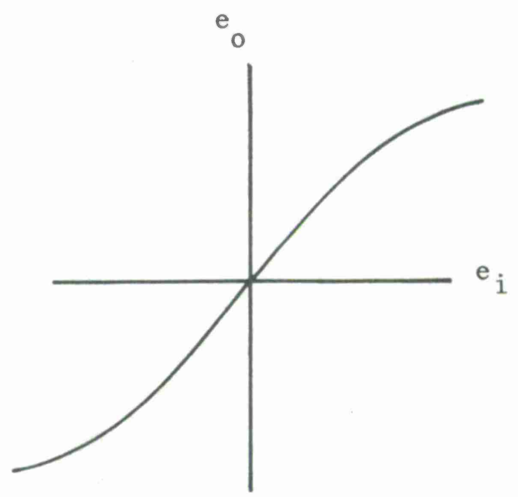


Fig. 6. Nonlinear voltage transfer characteristics of two-port.

TN-1975-65(7)

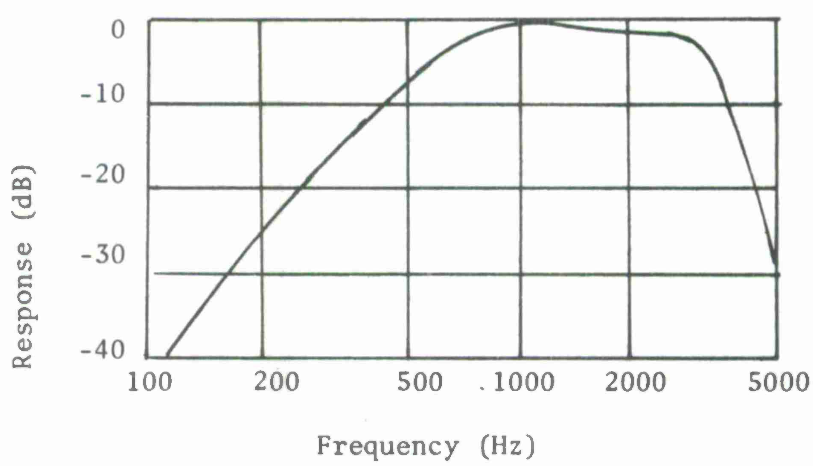


Fig. 7. C-message frequency weighting function.

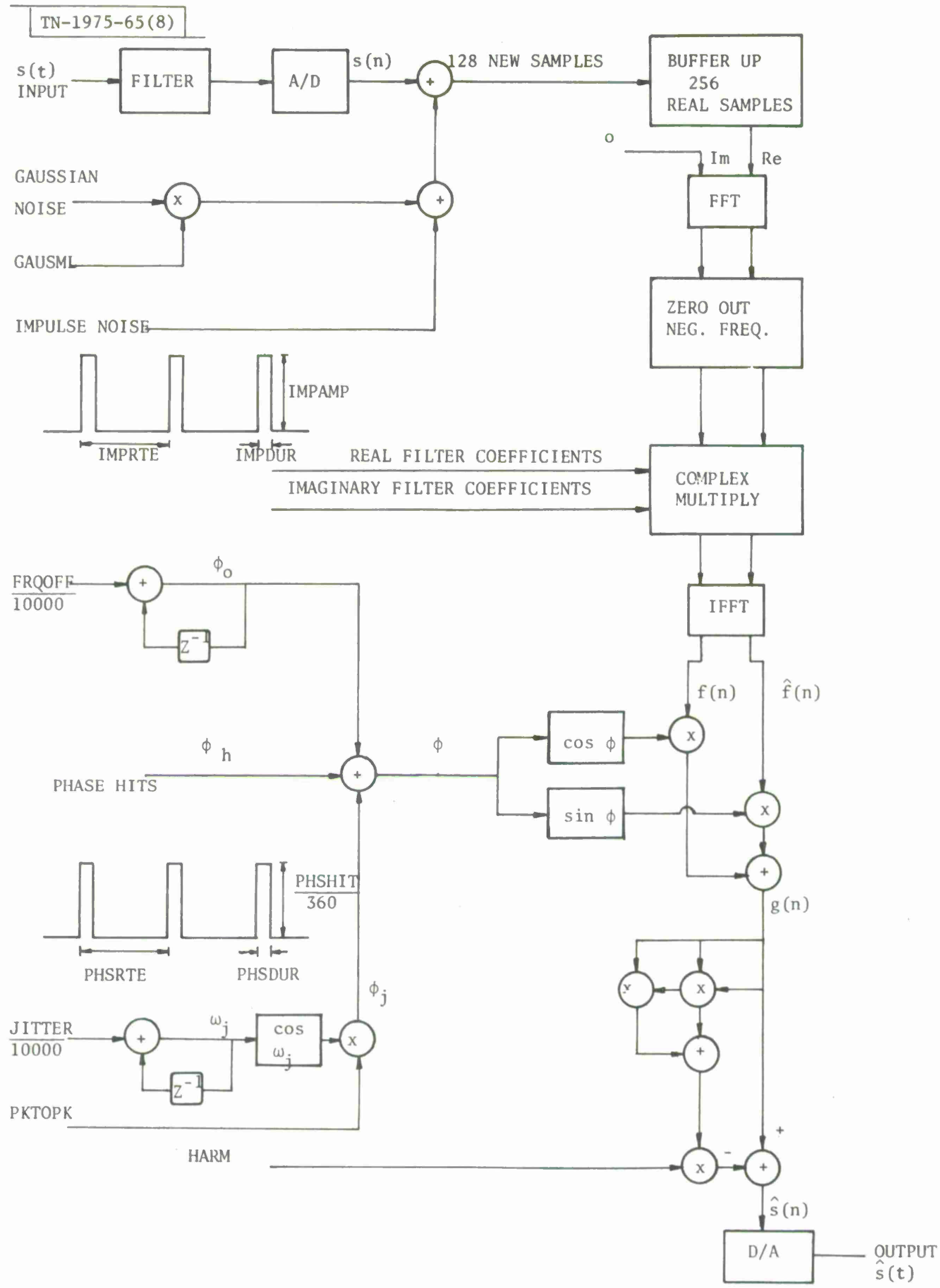
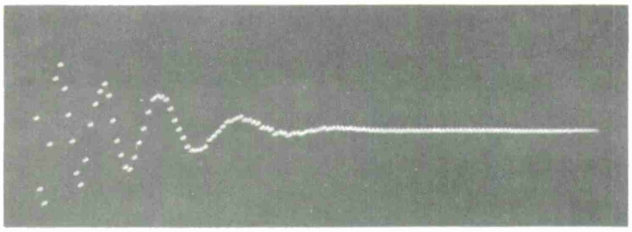


Fig. 8. Block diagram of telephone channel simulator.

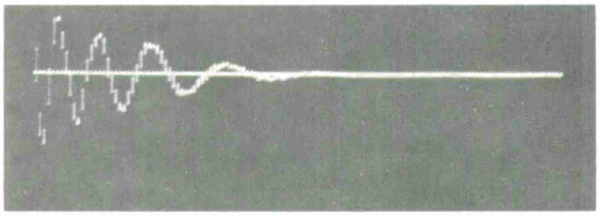
DATA FOR CONUS POOR VOICE

012403	003720	PHSRTE	2000.	:PERIOD BETWEEN PHASE HITS IN .01. sec UNITS
012404	000001	PHSDUR	1	:DURATION OF PHASE HITS IN .01. sec UNITS
012405	000001	FRQOFF	1	:FREQUENCY OFFSET IN HZ
012406	000017	PKTOPK	15.	:DEGREES PEAK TO PEAK OF JITTER
012407	000074	JITTER	60.	:HZ
012410	000021	PHSHIT	17.	:DEGREES
012411	050475	HARM	.6347.	:HARMONIC DISTORTION FACTOR
012412	001320	GAUSML	.022.	:NEW SIGMA = GAUSML*.25.
012413	000000	IMPRTE	0	:PERIOD BETWEEN IMPULSE HITS IN .01. sec UNITS
012414	000000	IMPAMP	.0.	:AMPLITUDE OF IMPULSE HITS
012415	000001	IMPDUR	1	:DURATION OF IMPULSE HITS IN .01. sec UNITS
012416	000000	PKTFRC	.0.	:FRACTIONAL PART OF PKTOPK
012417	000000	OFFFRC	.0.	:FRACTIONAL PART OF FRQOFF

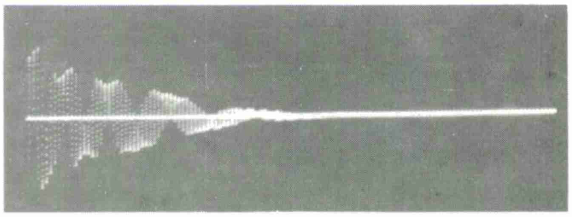
Fig. 9. Parameters of synthesizer reside in locations 403-417 of Md.



a) Unit Sample Response



b) Same as in (a), but with phase jitter at 17° peak-to-peak, 60 Hz



c) Same as in (a), but with frequency offset between transmitter cosine and receiver cosine of 1 Hz

Fig. 10. Time exposure of unit sample response of filter for conus poor voice line.

TABLE 1.
Impairments of the Eight Telephone Channels Simulated

CHANNEL SIMULATED	PHASE HITS	FREQ. OFFSET	PHASE JITTER	HARMONIC DISTORTION	GAUSSIAN NOISE	IMPULSE NOISE
CONUS POOR VOICE	45/15 min @ 17°	1 Hz	15°pk to pk 60 Hz	-20 dbmc	-40 dbmc	none
CONUS MID VOICE	15/15 min @17°	0 Hz	11.6° pk to pk 60 Hz	-25 dbmc	-46 dbmc	none
CONUS POOR DATA	45/15 min @17°	1 Hz	14° pk to pk 60 Hz	-20 dbmc	-37 dbmc	none
CONUS MID DATA	5/15 min @17°	0 Hz	11° pk to pk 60 Hz	-28 dbmc	-45 dbmc	none
EUROPEAN POOR VOICE	405/15 min @42°	7 Hz	35° pk to pk 50 Hz	-37 dbmc	-39 dbmc	225/15 min 74 dbrnc
EUROPEAN MID VOICE	45/15 min @32°	3.5 Hz	26° pk to pk 50 Hz	-43 dbmc	-45 dbmc	25/15 min 74 dbrnc
EUROPEAN POOR DATA	135/15 min @42°	6 Hz	35° pk to pk 50 Hz	-37 dbmc	-39 dbmc	225/15 min 74 dbrnc
EUROPEAN MID DATA	135/15 min @22°	2.7 Hz	18° pk to pk 50 Hz	-43 dbmc	-45 dbmc	25/15 min 74 dbrnc

TABLE 2.

Values used for the parameters of the simulator to correspond to the data in TABLE 1.

	CONUS POOR VOICE	CONUS MID VOICE	CONUS POOR DATA	CONUS MID DATA	EUROP. POOR VOICE	EUROP. MID VOICE	EUROP. POOR DATA	EUROP. MID DATA
PHSRTE (.01 sec units)	2000	6000	2000	18000	222	2000	667	667
PHSDUR (.01 sec units)	1*	1	1	1	1	1	1	1
FRQOFF (Hz)	1	0	1	0	7.	3.	6.	2.
PKTOPK (degrees)	15.	11.	14.	11.	35.	26.	35.	18.
JITTER (Hz)	60.	60.	60.	60.	50.	50.	50.	50.
PHSHIT (degrees)	17.	17.	17.	17.	42.	32.	42.	22.
HARM	.6348.	.3455.	.6348.	.2563.	.0928.	.0464.	.0928.	.0464.
GAUSML	.022.	.0085.	.0275.	.0098.	.0198.	.0082.	.0195.	.0082.
IMPRTE (.01 sec units)	0	0	0	0	400.	3600.	400.	3600.
IMPAMP	0	0	0	0	.11.	.11.	.11.	.11.
IMPDUR (.01 sec units)	1*	1	1	1	1	1	1	1
PKTFRC (degrees)	0	.6.	0	0	0	0	0	0
OFFFRC (Hz)	0	0	0	0	0	.5.	0	.7.

*PHSDUR & IMPDUR are variables of the system which, however, were fixed at .01. sec for these simulations since there was no information given.

TABLE 3.
List of Program Subroutines and Time and Program Memory
Required in Telephone Channel Simulator

SUBROUTINE	FUNCTION	PROG. MEM.	TIME
GTSPCH	Fetch input speech from Mx, store in Md	29	.078 msec
SCLUP	Scale data as block left maximally, twice	39	.422 msec
FFTFWD	128 point forward FFT	137	1.2 msec
EOLL	Bit reversed even odd separation. Last loop of 256 pt FFT	109	.28 msec
FLTMUL	Complex multiply by filter coefficients	29	.28 msec
SVEBUF	Save filtered spectrum in Mx	19	.08 msec
IFFT1	128 pt. IFFT to obtain even numbered samples	94	1.20 msec
GTBUF	Save $f(n)$, $\hat{f}(n)$, even numbered, in Mx Fetch filtered spectrum from Mx	44	.12 msec
IFFT2	128 pt. modified IFFT to obtain odd numbered samples	-(shared)	1.20 msec
FTCEVN	Fetch $f(n)$, $\hat{f}(n)$, evens, back from Mx	14	.39 msec
PHASIT	Compute phase, compute $g(n) = f(n) \cos\phi + \hat{f}(n) \sin\phi$ Add nonlinear distortion	121	.96 msec
ADJSCL	Change $\hat{s}(n)$ from block floating to fixed point	29	.046 msec
STRSPC	Store $\hat{s}(n)$ in Mx	16	.039 msec
ADINT	Interrupt routine. Service A/D-D/A Add Gaussian and Impulse Noise to $s(n)$ Update counters for phase hits and impulse hits	109	.46 msec
TOTAL		789 = 77%	6.755 msec = 53%

TABLE 4.
Buffers in Outboard Memoyr, Mx, needed for Telephone Simulation

BUFFER	ALLOCATION	BUFFER SIZE
EVIO	Even numbered in-out samples	64
ODIO	Odd numbered in-out samples	64
EVNEW	Most recent 64 even numbered samples of s(n)	64
ODNEW	Most recent 64 odd numbered samples of s(n)	64
EVOLD	Previous 64 even numbered s(n) } also as Mx Real	64
ODOLD	Previous 64 odd numbered s(n) } also as Mx Real	64
MXIMAG	Temp. storage of imaginary FFT output	128
GAUS	Midpoints of 256 equal areas of positive half of Gaussian	256
FLTR	Bit reversed filter coefficients (128 Real, 128 Imag.)	256
SINE	Sine of 0 to $\pi/2$ for FFT's and for computation of cosine ϕ	64
RVSINE	Bit reversed sine of 0 to $\pi/2$ for last stage of forward FFT	64
TOTAL		1152 = 56%

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digital vocoders

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

A telephone channel simulator has been implemented on the Lincoln Digital Voice Terminal (LDVT). This technical note first reviews the various distortions that occur in telephone channels and then describes the implementation. This work was undertaken at the suggestion of Mr. Ronald Sonderegger of the Defense Communications Engineering Center who observed that a real-time simulation on the LDVT should be far less costly than previous implementations. The results obtained indicate that the potential of the LDVT goes beyond the original intent of its use as a speech digitizer.

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